TR-1349

DOPPLER WAVE RECOGNITION WITH HIGH CLUTTER REJECTION
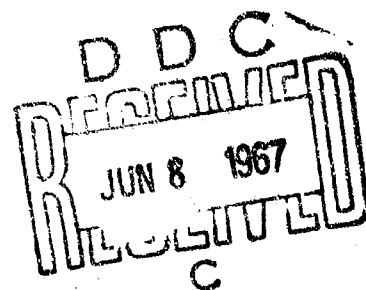
by

Dr. H. P. Kalmus

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U.S. ARMY MATERIEL COMMAND

HARRY DIAMOND LABORATORIES

WASHINGTON, D.C. 20438

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ABSTRACT

A Doppler Radar is described which makes it possible to detect
moving targets deeply embedded in clutter noise. The system is
based on the use of quadrature detection and a new low level cor-
relator. Most of the required amplification takes place after
correlation and integration so that amplifier overloading can be
avoided.

1. INTRODUCTION

The detection of moving targets is u~ lly based on the recogni-
tion of the Doppler wave. In its simplest form, a device for this
purpose consists of a CW transmitter and a receiver in which the
reflected wave is superimposed on part of the emitted wave. After
rectification, energy at the difference frequency, i.e., the Dop-
pler frequency, is obtained. Such devices have a poor signal-to-
noise ratio because any amplitude modulation occurring in the
transmitter or generated by fluctuating target return will cause
signals that can simulate a Doppler wave and possibly overload the
difference-frequency amplifier. Improvement is obtained by pro-
viding range sensitivity in the receivers by the use of pulse
techniques or frequency modulation. The clutter contributed by
nearby targets is thus attenuated. The clutter produced by far-
away targets, however, may still produce severe interference and
hamper recognition of the moving target.

There exists a pronounced difference between Doppler waves
and clutter waves: The Doppler energy from a single continuously
moving target appears as a continuous wave at either lower or
higher frequency compared with the frequency of the transmitted
energy. The energy produced by clutter, however, will be dis-
tributed on both sides of the transmitter frequency, especially if
observed over a certain period of time.

It was shown before[1] how, by the use of a quadrature detection
method, it becomes possible to distinguish between approaching and
receding targets. This was done without the use of critically
tuned filters, and led to a rugged system highly resistant to the
most severe environmental conditions.

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[1] Direction Sensitive Doppler Device. Proc. IRE, June 1955, pp. 698-
700; US Pats #2,934,756, 3,024,456.

It shall be shown now how a similar method in combination
with a novel correlator can be employed to reject clutter inter-
ference.

2. THE NEW SYSTEM

A signal $A \cos \omega_1 t$ is emitted and the wave $B \cos(\omega_2 t + \varphi)$ is
reflected. If $\varphi = 0$, the conditions are as shown in the phasor-
diagram of figure 1. For a receding target ω_2 is smaller than ω_1.
If an increasing phase angle is denoted by counterclockwise rota-
tion, B will rotate clockwise with an angular velocity $\omega_d = \omega_1 - \omega_2$.
After the signals are mixed in a square-law device, we obtain

$$(A \cos \omega_1 t + B \cos \omega_2 t)^2 = A^2 \cos^2 \omega_1 t + B^2 \cos^2 \omega_2 t + 2AB \cos \omega_1 t \cos \omega_2 t.$$

The last term, the only one of present importance, is equal to
$AB \cos (\omega_1 - \omega_2) t + AB \cos (\omega_1 + \omega_2) t$. The first term
$E = AB \cos \omega_d t$ contains the difference frequency and represents
the mixer output. Because it is a cosine term, the projection of
B on A along the time-axis supplies the shape of the Doppler wave.
It can easily be seen that the same Doppler wave is obtained
whether B rotates clockwise or counterclockwise. In other words,
the mixer output is the same for receding or approaching targets
and targets with reciprocating motion will be essentially indis-
tinguishable from targets with unidirectional motion.

Now, let us apply a second mixer in which the local signal
is delayed by $\pi/2$ and to which is fed the return wave in the same
manner as before. The B phasor is again vertical, but the A
phasor appears in the horizontal position (fig. 2). Compared with
figure 1, there is an important difference: For receding targets
the wave is delayed by $\pi/2$ and for approaching targets it is
advanced by $\pi/2$. Mathematically.

For receding targets: $\omega_{dr} = (\omega_1 - \omega_2) = \omega_d$. Denoting the
mixer output E',
$$E'_r = AB \cos (\omega_d t - \pi/2).$$

For approaching targets: $\omega_{da} = -\omega_d$.

$$E'_a = AB \cos (-\omega_d t - \pi/2) = AB \cos (\omega_d t + \pi/2).$$

If a correlator for comparing E with E' is available with output,
polarity positive for a phase shift of $+\pi/2$ between the two inputs,
and negative for $-\pi/2$, clutter can be suppressed by introducing
an appropriate integration network following the correlator.

An induction motor with its rotor held from turning by a
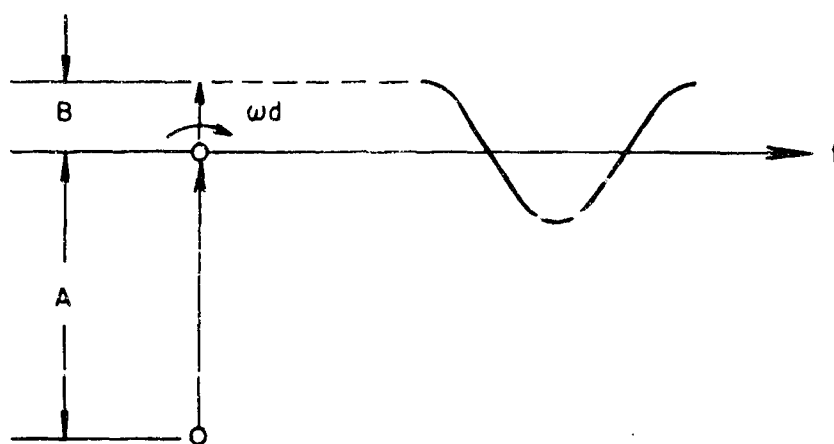barium titanate torque-measuring device so that positive voltage
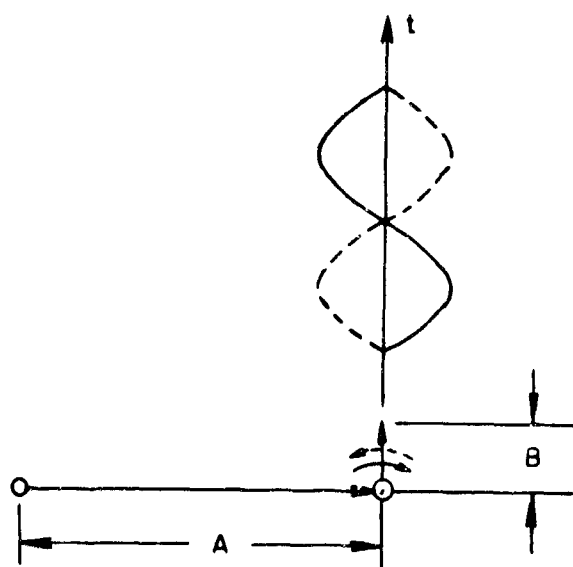
Figure 1.   Phasor diagram.

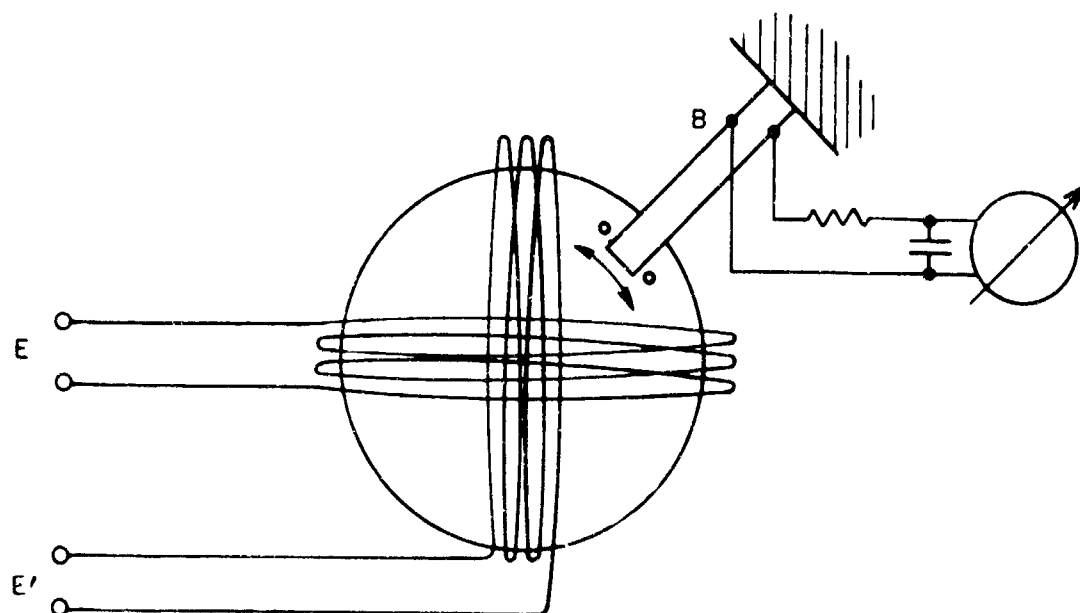Figure 2.   Phasor diagram, local signal delayed.

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Figure 3. Quadrature correlator.

is obtained for clockwise torque and negative for counterclockwise torque, as shown in figure 3, is such a correlator. The rotor may be an aluminum disc located in the field of two quadrature coils which are fed by E and E'. The output of the barium titanate element is fed through an RC integrator to a voltmeter. A moving target will produce a steady positive or negative integrator output. An oscillating target will produce an alternating voltage and give zero output if the integration time is sufficiently long. Amplitude modulation of A or A' will produce a zero or π phase shift between E and E', hence no torque. Random noise, being equivalent to AM and symmetrical FM of A and A', will also produce no output. The device is linear, i.e., noise waves in the coils with amplitudes many orders higher than the desi   .urrents do not alter the torque. This would not be the case if the disc could rotate; noise currents then would produce a braking action, thereby reduci   .'' ''.:nsitivity of the correlator.

Because of eddy-current losses, the efficiency of the disc-correlator is low. The device has been described because of its academic interest. A more efficient and more practical device will now be described.

As shown in figure 4, E and E' are produced as before, but the new correlator is fed by currents i_1 and i_2 derived from waves E and E'', E'' being derived from E' through a second $\pi/2$ delay network.

$$E'' = E'e^{-j\ \pi/2}$$

$$E''_r = AB \cos (\omega_d t - \pi)$$

$$E'_a = AB \cos \omega_d t.$$

These conditions are shown in figure 5.

Comparing the E wave with the E'' wave, we notice that for approaching targets the waves are in phase. For receding targets, they are in opposition. Hence, the correlator has to supply, say, a positive voltage for in-phase inputs and a negative voltage for out-of-phase inputs. Again, the operation must be linear; the output must be independent of noise waves orders of magnitudes higher than the desired information. In addition, the correlator must be a "real" multiplier; one wave alone should not produce any output. If this is not the case, we would measure in addition to the desired information the envelope noise of the individual noise waves fed into the correlator.
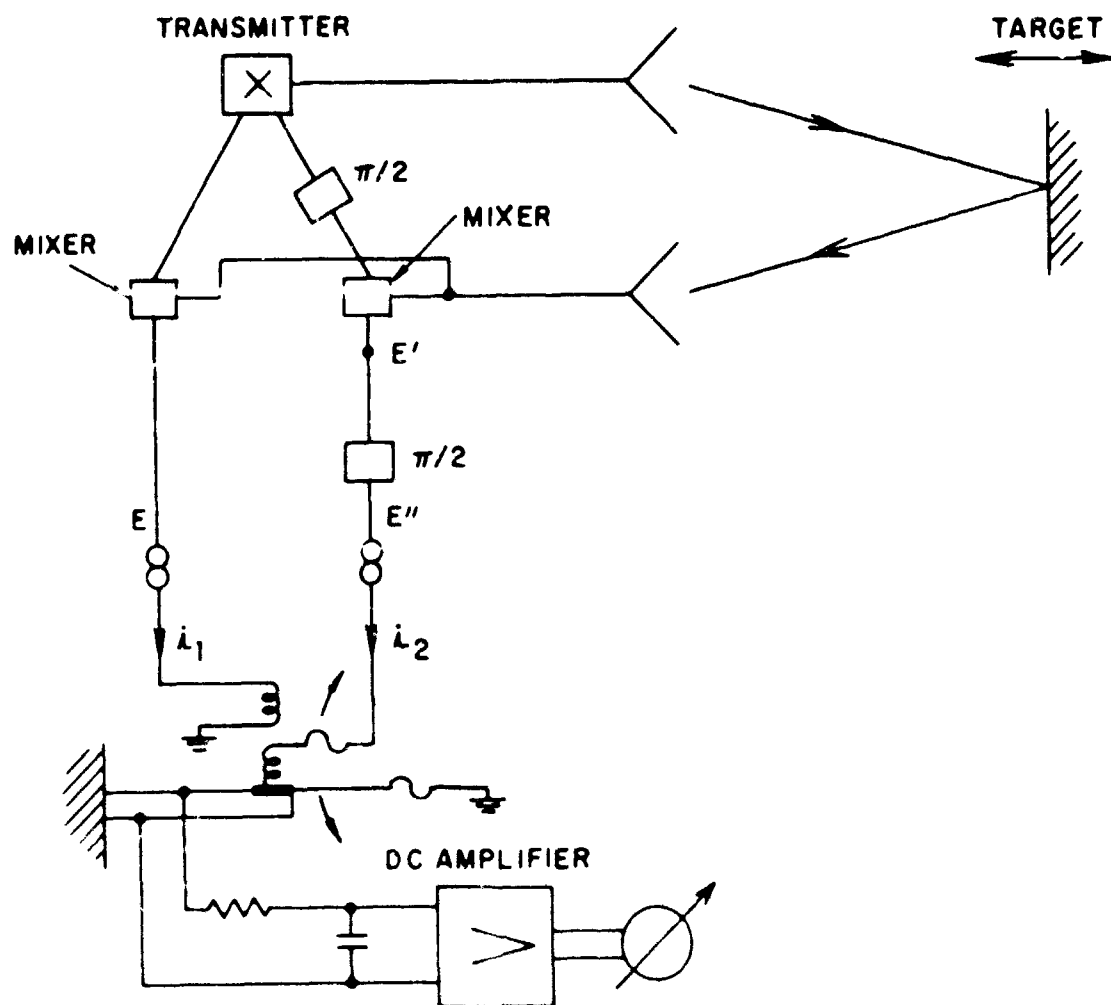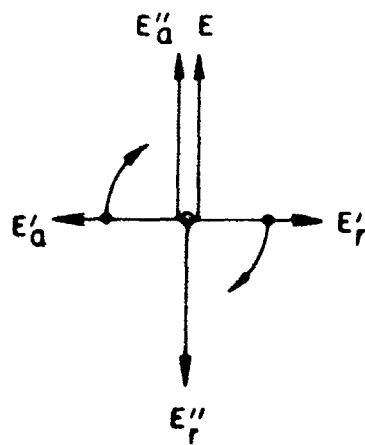
Figure 4. In-phase correlator.

Figure 5. The effect of the second delay network.

Ordinary phasemeters do not fulfill these conditions. Such devices contain electronic switches requiring one of the inputs to operate at a rather high level. Hence, amplifiers preceding the phasemeter are required and, due to overloading by high noise signals, the linearity may be lost. Neither is the second requirement fulfilled because, for a very small signal-to-noise ratio of both signals, balancing schemes offer no advantages. We need to find a correlator that is a <u>real</u> multiplier, working at a very low input level. <u>Any major amplification should take place after correlation and integration.</u>

In figure 4, there is shown schematically an electromechanical correlator that fulfills the above described requirements. It consists in its simplest form of two coils, one fixed and the other attached to a bimorph piezoelectric strip. If this strip is bent in one direction, it produces a positive, and if bent in the other direction, a negative voltage. The two coils are fed by the E and E" waves.

Denoting i_1 and i_2 the two currents through the two coils, M their mutual inductance, and s the axial excursion, of the moving coil the force producing bending of the ceramic strip is:

$$F = i_1 i_2 \frac{\partial M}{\partial s}.$$

Thus the device is linear only for small excursions. In other words, the transducer has to have a high mechanical impedance (it should be rather stiff). This requirement is fulfilled if the force is measured by a piezoelectric element. By the use of ferrite cores and by optimizing the form factor of the coils, $\frac{\partial M}{\partial s}$ can be made large and losses can be kept small so that a high sensitivity can be obtained. The two coils should preferably be driven from constant current generators such as pentodes or solid-state devices with small gain, so that overloading is avoided.

With reference to figure 4, the use of two phase shifters, one for the local high-frequency wave and the other for the mixer output, may not be understood. It may seem that a single phase shifter with twice the delay should serve the same purpose; this objection is based on the fact that in a superheterodyne receiver it is irrelevant whether a phase shift is introduced before or after the mixer since the result is the same. This is correct as long as the difference frequency i_d does not change its sign. In our case, however, i_d can be positive or negative. Hence, as shown before, the first phase shifter produces a delay of the mixer output for receding and an advance for approaching targets. The second phase shifter produces a delay, independent of target motion.

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3. IMPROVEMENTS

The simple scheme as shown in figure 4 has two drawbacks:

(1) The second phase shifter has to produce the $\pi/2$ delay over a rather wide range of frequencies.

(2) The correlator output is a small d-c voltage, and d-c amplification with high sensitivity presents difficulties. In addition, the correlator will also produce an interfering offset voltage.

The phase shifter difficulty can be avoided by replacing the single $\pi/2$ delay network in the E' branch by two $\pi/4$ networks, one producing a delay in the E' branch and the other producing an advance in the E branch. Denoting the output of the two phase shifters U' and U, and assuming that E' and E are in phase, it will be shown that U' and U have a phase difference of $\pi/2$ for all frequencies.

In figure 6, the the network is shown for E'.

$$ i = \frac{E'}{R + \frac{1}{j\omega C}} = E' \frac{j\omega C}{1 + j\omega RC}; \quad U' = \frac{1}{j\omega C} \, i = E' \frac{1}{1 + j\omega RC} \cdot $$

In figure 7, the network is shown for E.

$$ i = E \frac{j\omega C}{1 + j\omega F}; \quad U = Ri = E \frac{j\omega RC}{1 + j\omega RC} \cdot $$

Denoting the phase shifts for U' by φ_1 and for U by φ_2, we obtain:

$$ \tan \varphi_1 = -\omega RC $$

$$ \tan \varphi_2 = \frac{1}{\omega RC} \cdot $$

This trigonometric relationship is shown in figure 8 and it can be seen that $|\varphi_1 + \varphi_2| = \pi/2$. If, therefore, E' and E are out of phase by $\pm\pi/2$, U' and U will be in or out of phase for all target speeds. To maintain the signal-to-noise ratio for the device, it would be desirable to keep the product $\overline{UU'}$ constant for all frequencies. It will be shown that, although the frequency dependence is not actually zero, it is acceptably small over a suitably wide range of frequencies.
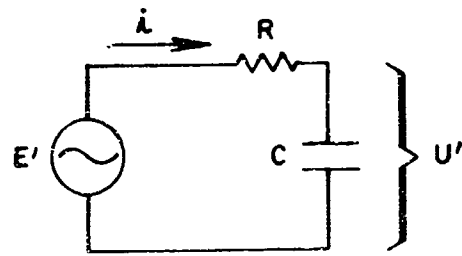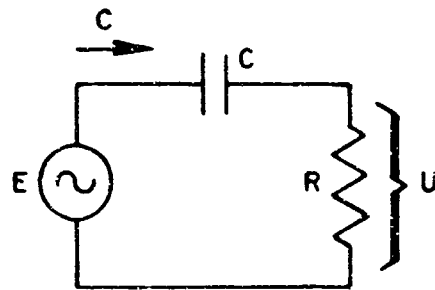
Figure 6. π/4 network.

Figure 7. π/4 network.

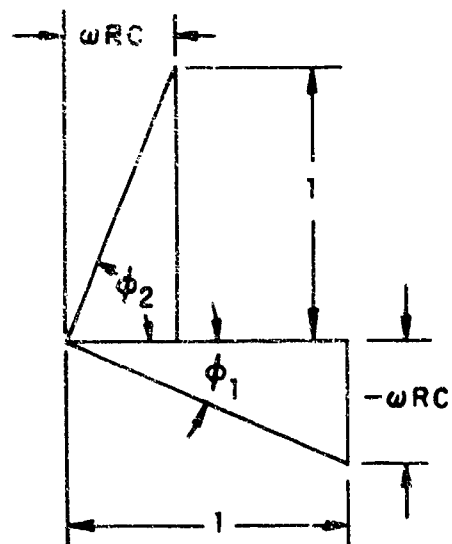Figure 8. Trigonometric relationship.

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$$\overline{U}' = \frac{1}{\sqrt{1 + \omega RC}} \quad ; \quad \overline{U} = \frac{\omega RC}{\sqrt{1 + \omega RC}} \; .$$

$$\overline{U}\,\overline{U}' = \frac{\omega RC}{1 + \omega RC} \; .$$

Now, if RC is chosen equal to $\dfrac{1}{\omega_0}$ (ω_0 being the most probable Doppler frequency), we can write:

$$\omega = \omega_0 + \Delta\omega \quad ; \quad \overline{U}\,\overline{U}' = \frac{(\omega_0 + \Delta\omega)\frac{1}{\omega_0}}{1 + (\omega_0 + \Delta\omega)\frac{1}{\omega_0}} = \frac{1 + \frac{\Delta\omega}{\omega_0}}{2 + \frac{\Delta\omega}{\omega_0}} \; .$$

for $\dfrac{\Delta\omega}{\omega_0} = -1$: $\overline{U}\,U' = 0$

for $\dfrac{\Delta\omega}{\omega_0} = -\frac{3}{4}$: $\overline{U}\,U' = 0.2$

for $\dfrac{\Delta\omega}{\omega_0} = -\frac{1}{2}$: $\overline{U}\,U' = 0.33$

for $\dfrac{\Delta\omega}{\omega_0} = -\frac{1}{4}$: $\overline{U}\,U' = 0.43$

for $\dfrac{\Delta\omega}{\omega_0} = 0$: $\overline{U}\,U' = 0.5$

for $\dfrac{\Delta\omega}{\omega_0} = \frac{1}{4}$: $\overline{U}\,U' = 0.56$

for $\dfrac{\Delta\omega}{\omega_0} = \frac{1}{2}$: $\overline{U}\,U' = 0.6$

for $\dfrac{\Delta\omega}{\omega_0} = \frac{3}{4}$: $\overline{U}\,U' = 0.64$

for $\dfrac{\Delta\omega}{\omega_0} = 1$: $\overline{U}\,U' = 0.67$

In figure 9, $\overline{U}\,\overline{U}'$ is plotted versus $\dfrac{\Delta\omega}{\omega_0}$ and it can be seen that the product does not show significant variations even for Doppler frequencies varying by a factor of 4. If the phase shifter is designed so that $\omega_0 = 1/RC$ and $\omega_0 = 100$ Hz, the product is 0.5 for 100 Hz, 0.33 for 50 Hz, and 0.67 for 200 Hz.

The second drawback of the simple scheme (fig. 4), namely the requirement of d-c amplification, can be avoided by replacing the first $\pi/2$ network by a phase shifter that will alternately produce
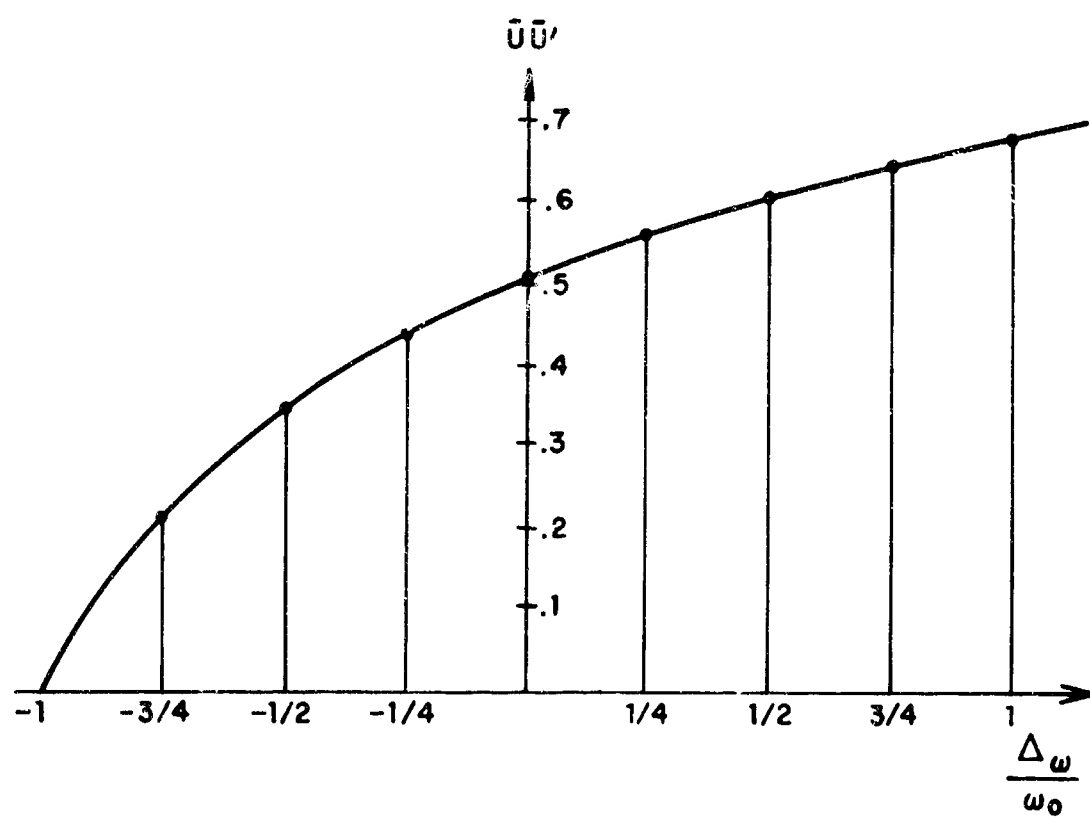
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Figure 9. The dependence of $\bar{U}\bar{U}'$ on frequency.
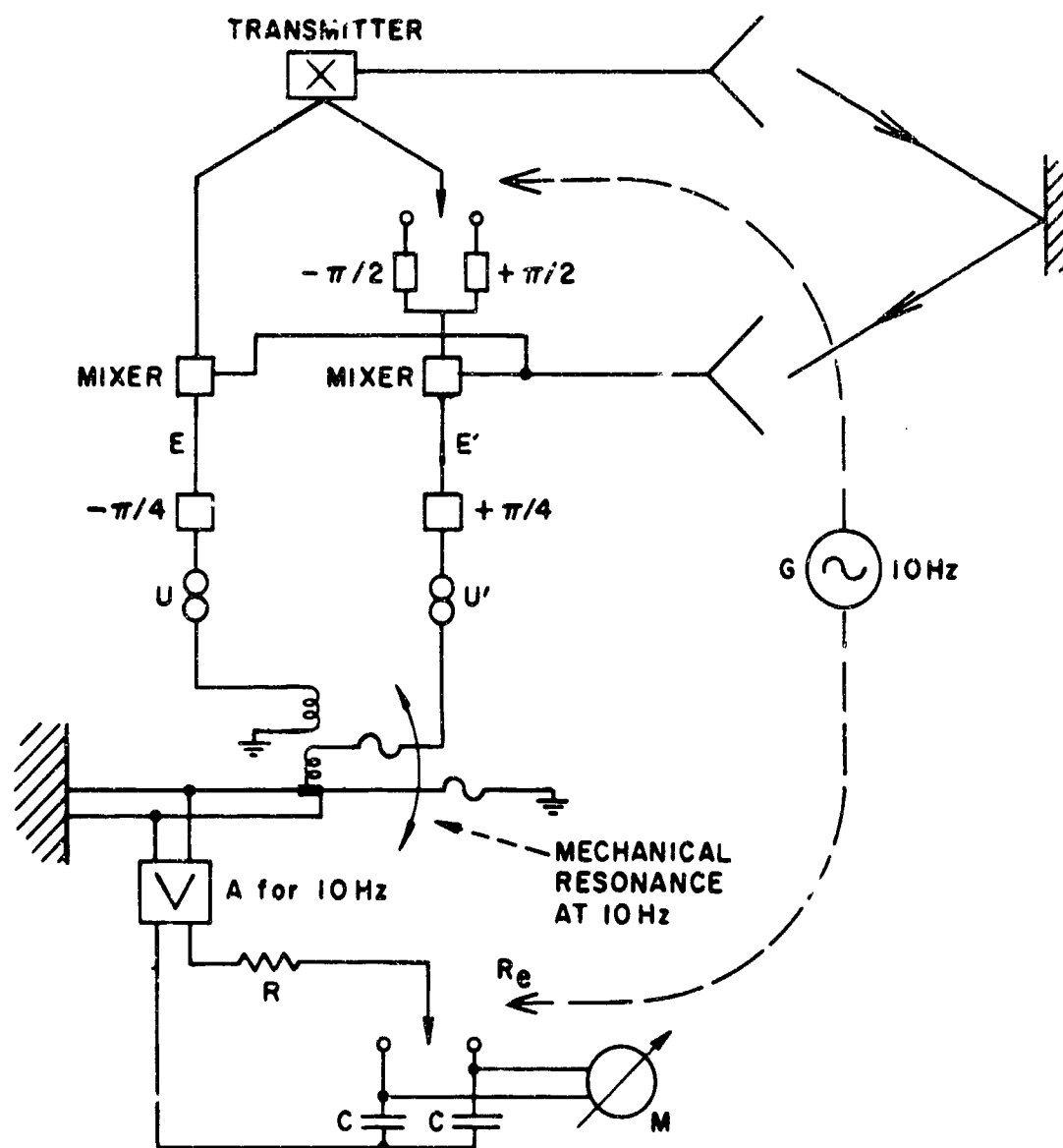
Figure 10. Correlator with switching.

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a shift of $+\pi/2$ and $-\pi/2$ at a predetermined rate smaller than the lowest expected Doppler frequency.

The improved scheme is shown in figure 10. The single high-frequency phase shifter of figure 4 is replaced by two networks inserted alternately by a switch driven from the 10-Hz generator G. The second phase shifter is replaced by the two $\pi/4$ networks as shown. The voltage U is now, as the voltage E before, a continuous Doppler wave. In the scheme of figure 4, E'' was a Doppler wave, with phase changing from 0 to π with respect to E for approaching and receding targets.

In the new scheme, U' is again a Doppler wave whose phase, however, alternates between 0 and π at the switching rate of 10 Hz as shown in figure 11. Assuming eight Doppler cycles for a complete switching period, figure 11a shows the condition for an approaching target. While the switch is at the left, E'' is in phase with E. While the switch is at the right, the voltages are in phase opposition. Hence, we obtain first attraction of the two coils and then repulsion. In figure 11b we see the conditions for a receding target. We obtain now first repulsion and then attraction.

The voltage delivered by the piezoelectric element is proportional to the force. It is a wave at switching frequency whose phase reverses with respect to the output from G if the target reverses its direction. Amplifier A is now a stable a-c amplifier with output detected in the phase-sensitive rectifier Re. There are now two means of integration: First, the piezoelectric element together with the movable coil is resonant at 10 Hz; second, electric integration is derived from the RC network feeding meter M.

If the a-c method according to figure 10 is chosen, it is necessary either to design a balanced structure of the correlator so that mechanical vibrations of the support do not create excessive noise signals in the 10-Hz frequency range, or to isolate the correlator acoustically from its surroundings.

4. EXPERIMENTAL RESULTS

In the Harry Diamond Laboratories, a figure-4 device has been built. With the use of x-band waves and two simple horns as antennas, small targets moving in one direction could be easily detected in the presence of clutter signals exceeding the target return by many orders of magnitude, although an electronic correlator was employed instead of the electromechanical correlator shown in figure 4, which ought to be much more effective.
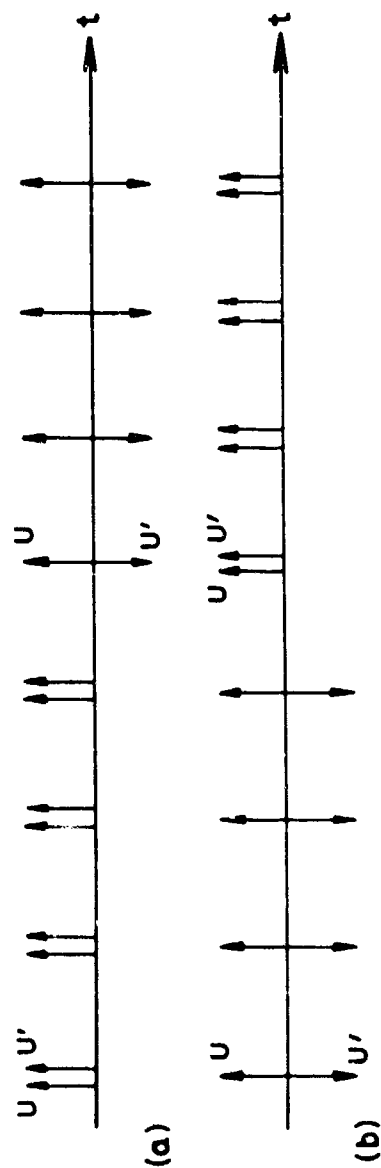
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Figure 11. Phase conditions of the doppler waves.

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11. SUPPLEMENTARY NOTES	12. SPONSORING MILITARY ACTIVITY
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13. ABSTRACT

A Doppler Radar is described which makes it possible to detect moving targets deeply embedded in clutter noise. The system is based on the use of quadrature detection and a new low level correlator. Most of the required amplification takes place after correlation and integration so that amplifier overloading can be avoided.

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Quadrature detector						
Correlator						